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REPORT

SATELLITE BROADCASTING :

differential demodulation of a 2-4 PSK signal using a three-port SAW delay line

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Section **Page**

Summary

1. Introduction 1
2. Differential demodulation 1
3. The SAW delay line 1
4. Experimental work 2
5. Results 3
6. Conclusions 6
7. Acknowledgement 6
8. References 7

Appendix 8

It uses SI units in accordance with B.S. document
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Summary

This Report describes a SAW (Surface Acoustic Wave) device which can be used to differentially demodulate the 2-4 PSK (Phase Shift Keying) data signal in a C-MAC receiver. It shows that the system can be optimised so that the performance is good even in the presence of typical instrumental inaccuracies caused by manufacturing tolerances and thermal expansion.

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1. INTRODUCTION

The European Broadcasting Union has proposed the use of the C-MAC system as a standard for direct broadcasting by satellite (DBS)[1]. In the C-MAC system, the f.m. chrominance and luminance signals are time-division-multiplexed with a 2-4 PSK (Phase Shift Keying) digital sound/data burst[2]. In 2-4 PSK modulation, the binary data is differentially encoded such that a data '1' and a data '0' correspond to a ninety degree phase advance and retardation respectively. This system has the advantage that it can be demodulated by coherent, differentially coherent and discriminator demodulators, allowing flexibility in the receiver design. Of these, the differentially coherent demodulator is a useful compromise as it offers good performance whilst being simple to implement; it does not require carrier recovery.

In a differential demodulator, a delayed version of the signal is used as a reference for demodulation. Conventionally, the delay is generated using a length of coaxial cable. However, this is both expensive and bulky for use in a domestic receiver, (about 10 metres of cable are needed for a 20 Mbit/s system) and is difficult to cut accurately on a production line.

An alternative is to use a SAW[3] device where the velocity of propagation is low enough for the required delay to be obtained using a substrate a few millimetres long. In addition, the intrinsic band-pass filter response of a SAW interdigital transducer (IDT) can be used as the receive filter for the demodulator.

This Report describes a differential demodulator based on a three-port SAW delay line. It shows how the system can be optimised so that the performance remains high even when manufacturing and operational inaccuracies are present.

2. DIFFERENTIAL DEMODULATION

Instead of demodulation using a locally generated carrier as in coherent demodulation, differential demodulation uses a delayed version of the original signal. A typical configuration is shown in Fig. 1. The signal is filtered to optimise the signal-to-noise ratio and then split into two paths. One path includes a one bit-period delay, after which the two signals are multiplied in a linear mixer. A low-pass filter removes any unwanted mixing components.

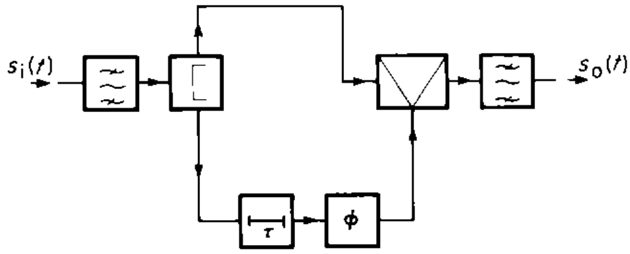


Fig. 1 - Differential demodulation.

For optimum performance of a demodulator of the type shown in Fig. 1, the carrier frequency f_o, delay T and phase shift ϕ are related by (see the Appendix):

$$2\pi f_o T + \phi = (2n + 1)\ \pi/2 \text{ (radians)} \quad \ldots (1)$$

Hence the choice of carrier is determined by the bit period and phase shift. The delay of one bit period τ, and the phase shift ϕ can be implemented separately. However, it is more convenient to simply extend the delay line slightly to give the required overall phase relationship. For example, operation of the demodulator with a 70 MHz centre frequency at a bit rate of 20.25 Mbit/s requires a total delay of 53.6 ns.

An important point to note from equation 1 is that an error in ϕ can be corrected by adjusting the carrier frequency to maintain the correct phase relationship. This is important when considering commercial production of this type of demodulator as it provides a method of correcting for manufacturing errors to achieve optimum performance. From the Appendix it can be shown that the output of the demodulator for an unmodulated carrier signal is:

$$V = \cos(2\pi f_o T + \phi) \quad \ldots\ldots\ldots\ldots\ldots (2)$$

This expression (V) is zero for the optimum phase relationship given by equation 1. Hence, this could be used as the basis of an automatic frequency control loop as the C-MAC specification[2] includes one line of reference carrier per television picture.

3. THE SAW DELAY LINE

A simple SAW interdigital transducer (IDT)

launches an acoustic wave in two opposite directions. This bidirectionality was utilised in this work to make a delay line with two outputs. The design of the delay line is shown in Fig. 2. A relative delay was obtained by positioning one of the output IDTs further from the input than the other. It can be seen that the two output IDTs were half the height of the input IDT, and were offset with respect to each other. This was necessary to prevent any interaction between the two outputs due to acoustic reflections. However, this increased the insertion loss by 6 dB, as only half the acoustic energy propagating in each direction was intercepted by the output.

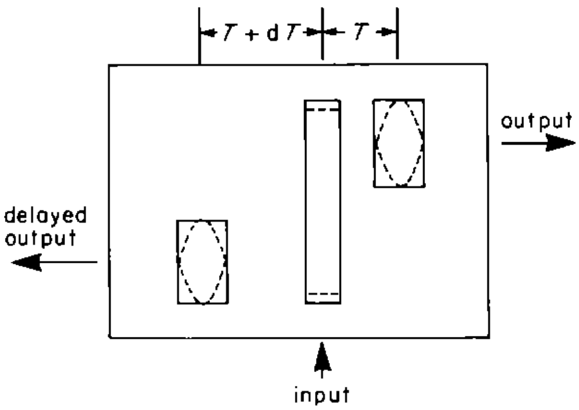


Fig. 2 - SAW delay line.

A possible problem with using a SAW device is the accuracy with which the relative delay can be produced. The accuracy depends on the errors in the mask manufacturing process, and the temperature dependence of the substrate material. IDTs can be positioned with relative accuracy of ± 2 μm on the experimental mask-making facility developed at the BBC's Research Department. For the YZ-cut lithium niobate substrate used in this work, this delay error corresponds to $\pm$ 0.6 ns. If the carrier frequency is 70 MHz, this error is equivalent to a phase error of $\pm 15^\circ$.

The second source of error in the relative delay is the temperature dependence of the substrate. Ideally, ST-cut quartz should be used as it has a temperature coefficient of delay which is sufficiently small for the application[3]. However, it has a low piezoelectric figure of merit which means that the insertion loss for a wideband filter of the fractional bandwidth required for this work (35%) would be unacceptably high (about 40 dB). For this reason, YZ-cut lithium niobate was used[4], which has a temperature coefficient of delay of 9×10^{-5} °C^{-1}. Hence a temperature variation of $\pm$ 25°C would give a phase error of $\pm 2.5^\circ$ for a 70 MHz carrier frequency.

Overall, it was expected that the dominant source of error would be due to mask manufacturing tolerances. In the delay line design where there are three IDTs, the expected possible phase error is $\pm 30^\circ$. An error of this magnitude would give a signal-to-noise ratio degradation of 1.3 dB. However, as explained earlier the demodulator performance could be optimised by tuning the carrier frequency.

The differential demodulation of 2-4 PSK has been studied using a computer simulation. The simulation uses complex arrays to represent the signal in the time and frequency domains. The model incorporates the measured filter amplitude and phase responses of all the major components in the transmission chain. In addition, the effect of echoes, and amplitude imbalances in the phase modulator were simulated. Overall, the simulation was found to agree with experiment to within 0.2 dB.

Using the simulation, it was found that a near-optimum receive filter pass-band shape was a Butterworth-type response with the 3 dB bandwidth equal to the bit rate. However, as explained in Section 2, it was necessary for the carrier frequency to be tuned to compensate for manufacturing errors. This meant that the bandwidth had to be increased to minimise performance degradation due to clipping the offset spectrum. The simulation showed that a receive filter bandwidth of 1.15 times the bit rate allowed the carrier to be tuned by $\pm$ 1.5 MHz for a degradation of less than 0.2 dB.

4. EXPERIMENTAL WORK

A differential demodulator incorporating a SAW filter delay was built. The circuit is shown schematically in Fig. 3.

A matching inductor was used in the input of the SAW delay line to reduce the insertion loss to about 25 dB. The insertion loss could have been reduced further by using inductors on the outputs as well, but this was not done to avoid differential phase-shifts in the two demodulator arms.

The two output signals from the SAW delay line were amplified and fed into a linear mixer. A low-pass filter was used to remove the second harmonic of the carrier frequency. The baseband data was then amplified before sampling.

The performance of the demodulator was measured using a 2-4 PSK modulator whose carrier frequency could be varied by tuning the local oscillator input to the phase modulator. A 27 MHz

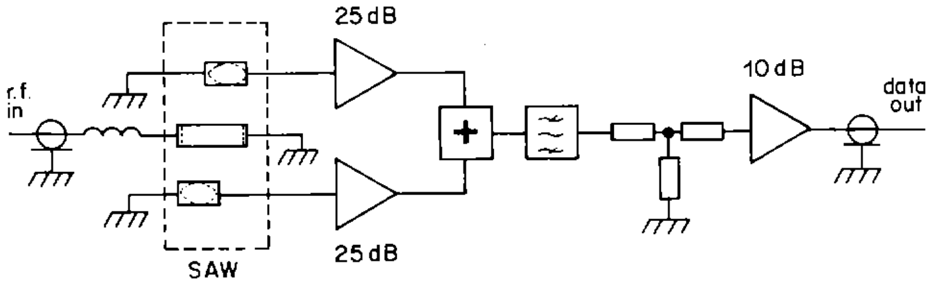


Fig. 3 - Circuit diagram of differential demodulator.

noise bandwidth transmit filter was used to shape the modulator output spectrum.

A broadband Gaussian noise source was used for the bit error ratio (BER) measurements. The power spectrum of the noise source was flat to within ± 0.1 dB over the range ± 20 MHz centred on 70 MHz. The signal power was measured after the transmit filter, and the noise power was measured in a 27 MHz noise bandwidth.

5. RESULTS

The demodulator was tested with three SAW delay lines, each with a different filter response. These are referred to as SAW 2, 3 and 4. (SAW 1 was a failure). In addition, a conventional cable-delay demodulator with a conventional Butterworth filter was tested for comparison. The measured frequency responses of the three SAW devices and the Butterworth filter are shown in Fig. 4. It should be noted that the figure does not show the 0.5 dB peak-to-peak ripple in the SAW filter pass-bands due to triple-transit echoes. Figs. 5(a), (b) and (c) show the frequency and group delay response of SAW 4 in more detail, with the ripple due to the echoes clearly evident. The computer simulation showed that echoes of this magnitude caused a degradation in performance of 0.1 dB. The first delay line tested in the demodulator, SAW 2, had a 3 dB bandwidth of 19 MHz. The later delay lines SAW 3 and SAW 4, had 3 dB bandwidths of 23.5 MHz to allow for tuning of the carrier frequency to optimise performance (see Section 3). The difference between SAW 3 and SAW 4 was a correction to the filter response to remove a ± 1 dB rise at the pass-band edge of SAW 3.

In testing each delay line, the carrier frequency was adjusted to give optimum performance of the demodulator. Fig. 6 shows the measured BER versus carrier frequency for the SAW 4 delay line. The figure also shows the d.c. voltage generated for an unmodulated carier. It

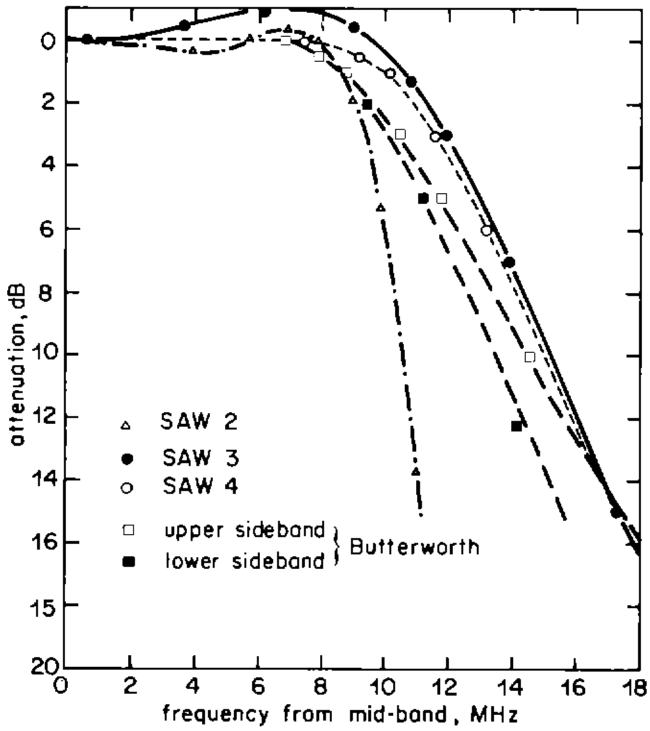


Fig. 4 - Responses of the experimental filters.

can be seen that the minimum BER was measured when the d.c. signal was close to zero, as predicted by equation 2. Hence this demonstrated the possibility of incorporating an AFC loop in the demodulator.

Fig. 7 shows the eye diagram of the demodulated data sequence using the SAW 4 delay line with the carrier set to optimum frequency. The well-defined eye indicates that the filter characteristic was near-optimum. The slight asymmetry of the eye was due to the centre frequency of the filter pass-band (69 MHz) not coinciding with the optimum carrier frequency (70.65 MHz).

Fig. 8 shows the experimental and simulated BER measurements for the different delay line demodulators. Also shown is the theoretical

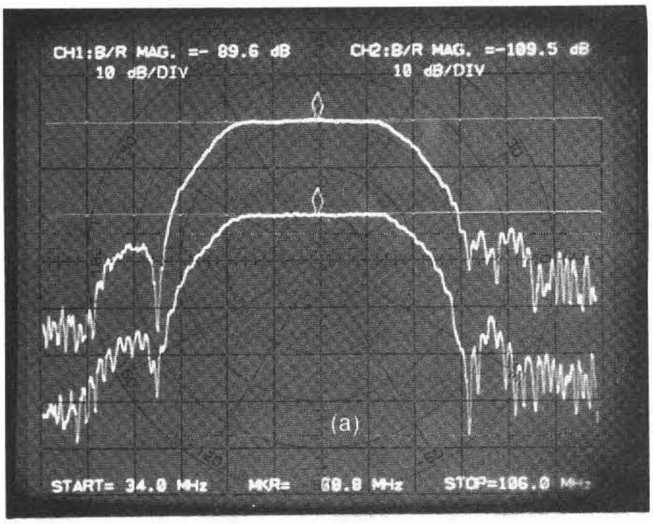
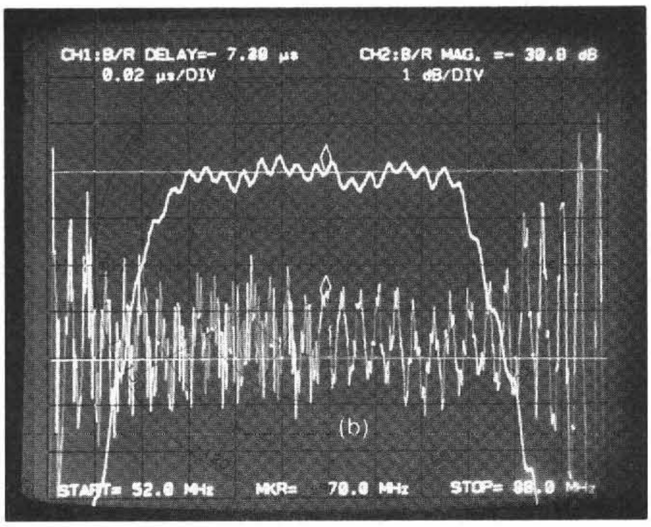


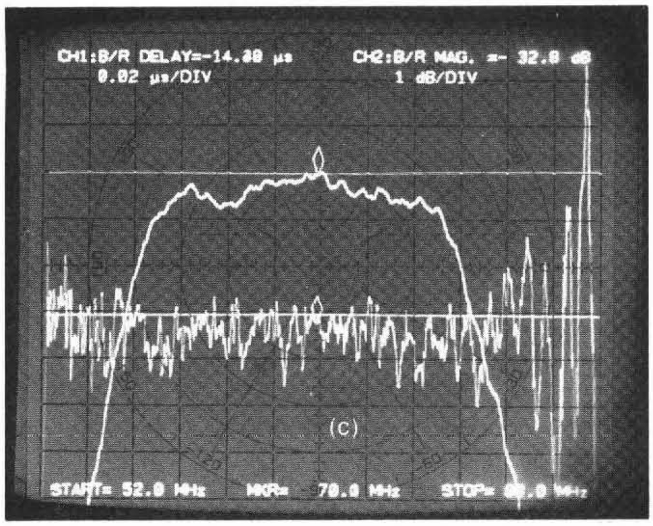
Fig. 5 - Frequency responses of filter SAW 4.

(a) Both outputs
Vertical scale : 10 dB/div.
Horizontal scale : 7.2 MHz/div.

(b) Output 1
Vertical scale : 1 dB/div (upper) and 20 ns/div (lower).
Horizonal scale : 3.6 MHz/div.

(c) Output 2
Vertical scale : 1 dB/div (upper) 20 ns/div (lower).
Horizontal scale : 3.6 MHz/div.





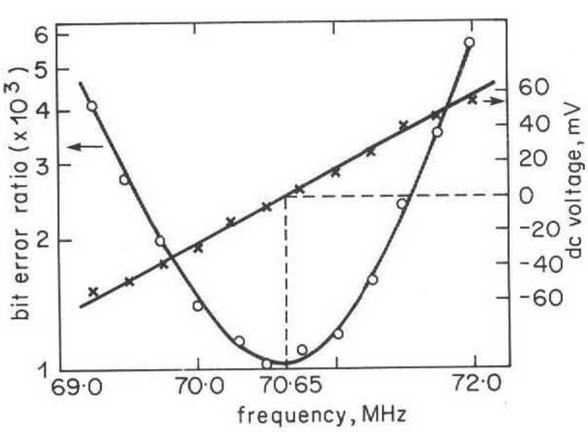


Fig. 6 - Variation of bit-error ratio and voltage for automatic frequency control with frequency.

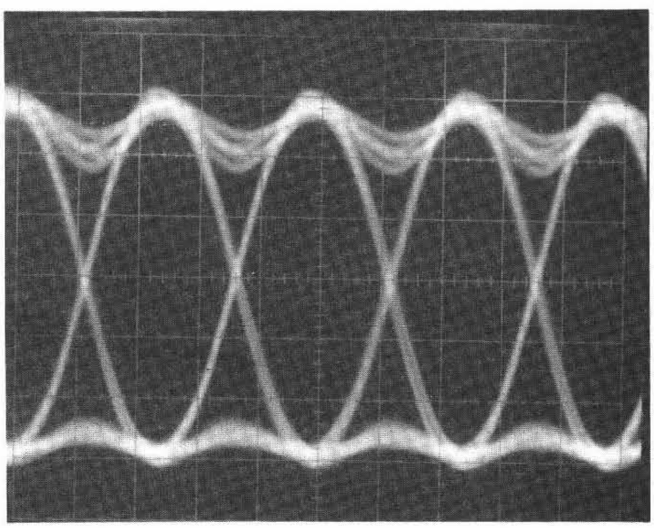


Fig. 7 - Data eye diagram for SAW 4 delay line.

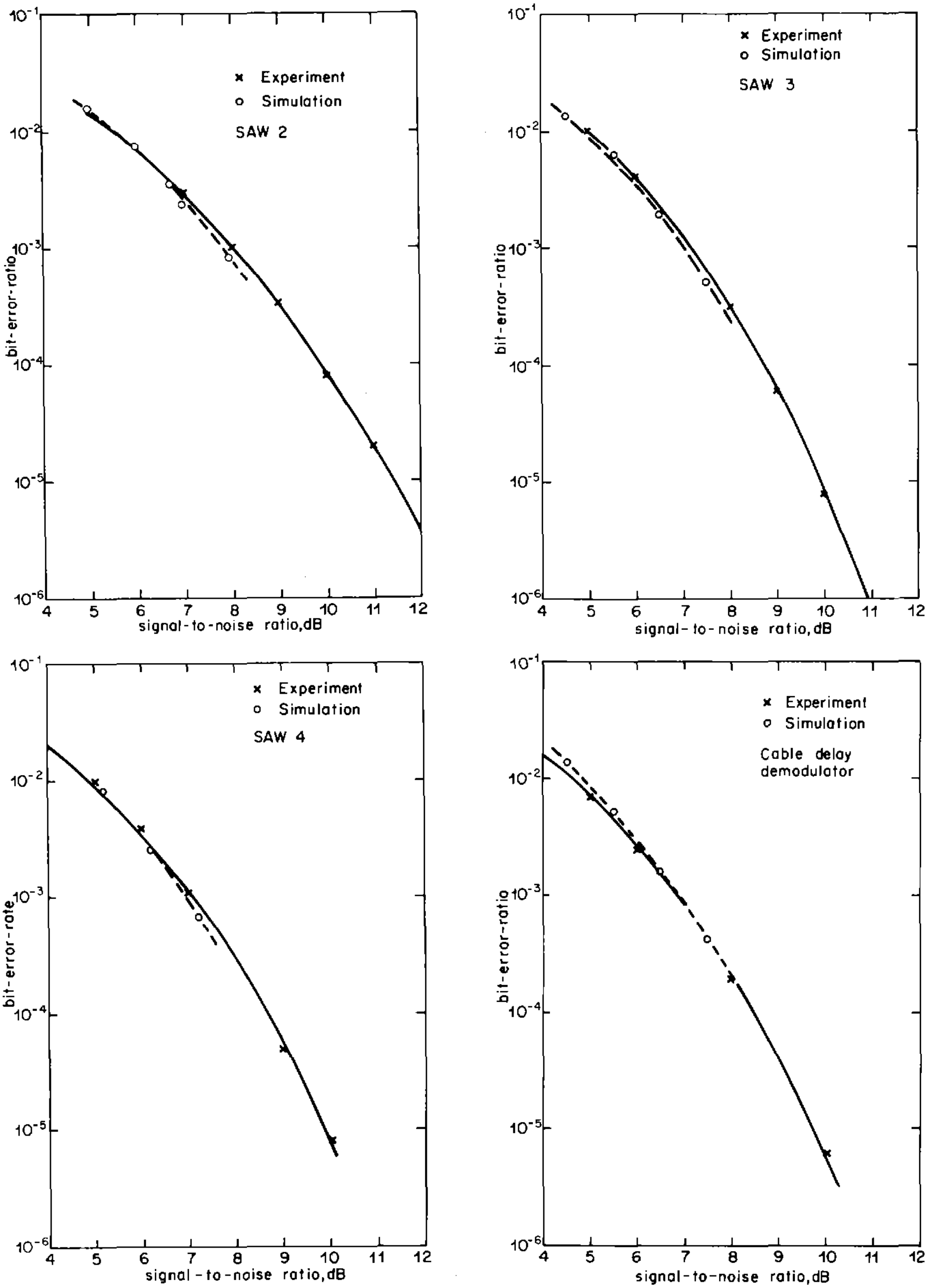


Fig. 8 - Bit-error ratio as a function of signal-to-noise ratio.

TABLE 1

S/N (dB) Required for a BER of 10^{-3}

The theoretical value is 6.7 dB for optimum conditions

	Experiment	Simulation	Bandwidth	
			MHz	Fraction of bit-rate
SAW 2	8.0 ± 0.1	7.8 ± 0.1	19.0	0.95
SAW 3	7.2 ± 0.1	7.1 ± 0.1	23.5	1.15
SAW 4	7.1 ± 0.1	7.0 ± 0.1	23.5	1.15
Cable	6.8 ± 0.1	6.8 ± 0.1	20.25	1.0

performance of an optimum 2-4 PSK channel. Table 1 summarises the results in terms of the signal-to-noise ratio required for a BER of 10^{-3}.

The agreement between experiment and simulation was good, generally to within 0.2 dB.

However, it was found that for high BERs the measured results were apparently better than the simulated results.

It can be seen from Table 1 that the best SAW demodulator (SAW 4) was within 0.4 dB of the theoretical value while the cable demodulator was within 0.1 dB. The difference was mainly due to the slightly wider bandwidth of the SAW filter to allow for carrier frequency tuning.

The performance of the narrow bandwidth SAW delay line, SAW 2, was 1 dB worse than the cable delay demodulation using the Butterworth filter. This degradation was mainly due to the rapid roll-off of the SAW frequency response, The computer simulation showed that a Butterworth filter with a bandwidth of 19 MHz would cause less degradation.

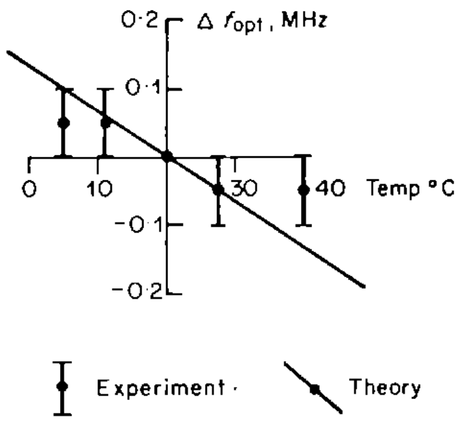


Fig. 9 - Variation of optimum centre frequency with temperature.

The variation of optimum carrier frequency with temperature was measured using an environmental test chamber. The results are shown in Fig. 9. The agreement with the expected values was reasonable, although deviations from linearity at the highest and lowest temperatures was probably due to the device not being completely in thermal equilibrium with the test chamber.

6. CONCLUSIONS

Differential demodulation of 2-4 PSK signals has been demonstrated using SAW delay lines. It has been shown that errors in manufacturing the SAW devices can be compensated for by varying the carrier frequency using an automatic frequency control loop, thus minimsing the received bit error ratio.

Three SAW delay lines were tested, with the final version working within 0.4 dB of the theoretical value for optimum demodulation. The degradation was mainly due to the need for a slightly wider-than-optimum bandwidth to allow for tuning of the carrier frequency. If the manufacturing errors were reduced, the bandwidth of the SAW device could be reduced to bring the performance closer to the theoretical value.

As SAW devices are ideally suited to mass-production, SAW delay lines of the type described in this Report could form the basis of domestic differential demodulators for satellite television.

7. ACKNOWLEDGEMENT

The authors wish to thank Dr. John Zubrzycki for making the SAW devices.

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Appendix

The relationship between the frequency delay and phase shift in the differential demodulator

Let the signal at the input of the demodulator be given by

$$s_i(t) = \cos[\,2\pi f_0 t + \theta(t)] \qquad \text{(A - 1)}$$

The phase function $\theta(t)$ carries the modulation information. It will advance or retard by $\pm\pi/2$ radians each bit interval τ depending on the modulation.

The delay line is such that the delay T is approximately equal to the bit interval τ.

Thus the signal at the output of the delay line is

$$s_i(t-T) = \cos[2\pi f_0 (t-T) + \theta(t-T) - \phi] \qquad \text{(A - 2)}$$

After mixing, the wanted signal is the baseband component of

$$y(t) = s_i(t) \,.\, s_i(t-T)$$

$$\text{i.e.}\quad s_0(t) = \cos[\,2\pi f_0 T + \theta(t) - \theta(t-T) + \phi] \qquad \text{(A - 3)}$$

For 2-4 PSK, the signalling is such that the phase advances or retards by $\pi/2$ each signalling interval. Although the phase change is notionally instantaneous, filtering will smooth the transition. However, it can be assumed that provided the filtering is not very tight and that the delay T is close to the bit period τ, then the signalling term in equation A-3 reduces to

$$\psi = 2\pi f_0 T + \theta(t) - \theta(t-T) + \phi \qquad \text{(A - 4a)}$$

$$= 2\pi f_0 T \pm \pi/2 + \phi \qquad \text{(A - 4b)}$$

To give a suitable signal for demodulation $\cos\psi$ should have the value ± 1 at the sampling instants, i.e.

$$2\pi f_0 T + \phi = \pm n\pi \qquad n = 0,1,2 \ldots \qquad \text{(A - 5)}$$

Putting (A - 5) in (A - 4b) gives

$$2\pi f_0 T + \phi = (2n + 1)\pi/2. \qquad \text{(A - 6)}$$

This is the desired relationship between the delay, T, the centre frequency, f_0, and the phase shift of the circuit, ϕ.

In the event that there is some inaccuracy in the phase shift or delay, some compensation can be obtained by modifying the centre frequency. The extent to which compensation for time delay errors is possible is limited by the requirement that the phase at the sampling times is $\pm\,\pi/2$ the change on the previous time.

A feedback signal can be obtained if plain carrier is present. In such a case

$$\theta(t) - \theta(t-T) = 0$$

Thus from (A - 4a) and (A - 3)

$$s_0(t) = \cos(2\pi f_0 T + \phi) \qquad \text{(A - 7)}$$

For correct operation, (A - 6) shows

$$s_0(t) = 0.$$

If there is any variation in f_0, T or ϕ, then $s_0(t)$ changes and can be used as a feedback signal (referred to as V in the main text).

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